RCA

LB-958

AN EXPERIMENTAL AUTOMOBILE RECEIVER EMPLOYING TRANSISTORS

RADIO CORPORATION OF AMERICA
RCA LABORATORIES DIVISION
INDUSTRY SERVICE LABORATORY

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An Experimental Automobile Receiver Employing Transistors

Approved

Stuart Wm. Seeley

An Experimental Automobile Receiver Employing Transistors

Introduction

This bulletin describes an experimental automobile broadcast receiver utilizing nine experimental junction transistors in a superheterodyne circuit. The receiver operates directly from the six-volt storage battery without vibrator, power transformer, or rectifier. The average current drain, including that for two pilot lights, is approximately one-tenth that of a conventional automobile receiver.

The performance of this receiver is comparable to that of conventional automobile receivers. Particular emphasis has been placed on maintaining performance over a wide range of ambient temperature, both to accommodate the severe requirements specified for automobile service, and to establish the operability over such a temperature range of apparatus employing *germanium* transistors.

The receiver circuits and performance characteristics, including performance data for the ambient temperature range -40 degrees C to +80 degrees C, are described in detail. Techniques which render circuit operation insensitive to variation of ambient temperature and which permit interchangeability of transistors are discussed.

General Description

The automobile receiver, shown in Fig. 1, uses nine experimental p-n-p alloy junction transistors in a superheterodyne circuit employing a 455-kc intermediate frequency.

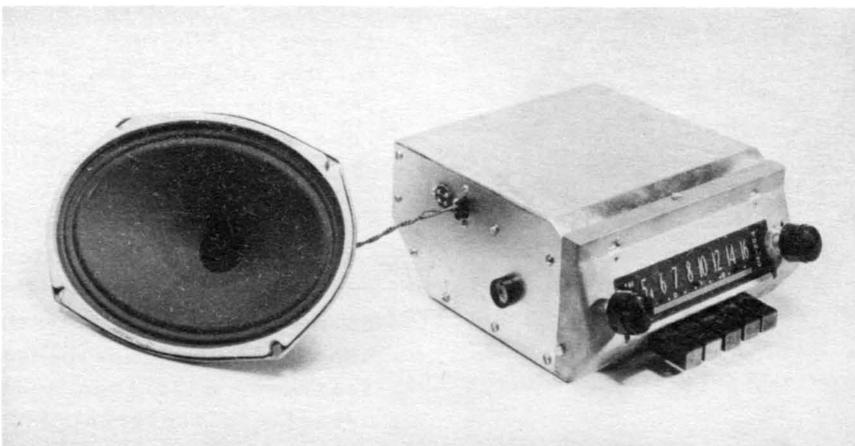


Fig. 1 - Experimental transistor automobile receiver.

A permeability-tuned r-f stage and a class-B transformer-coupled output stage are incorporated. The current drain of the radio is 250 ma for the zero signal condition; an additional 300 ma are required by the two pilot lights. On a sustained tone, with maximum power output the total drain rises to one ampere.

The performance characteristics of the receiver at 20 degrees C are summarized below:*

Sensitivity	2	μvolts
Noise Performance		
Input for 20 db S/N	12	μvolts
ENSI	0.4	μvolt
Power Output	2	watts
Selectivity (ACA)	41	db
AGC (Figure of Merit)	63	db

The operating temperature range of the receiver is -40 degrees C to +80 degrees C. Circuit techniques provide stabilization for the most part, and thermistors are used for temperature compensation in the audio amplifier.

The receiver is constructed in three main sections as shown in Fig. 2. These sections are the tuner assembly, in the foreground; the audio amplifier, mounted over the tuner; and the high frequency part of the receiver, in the background.

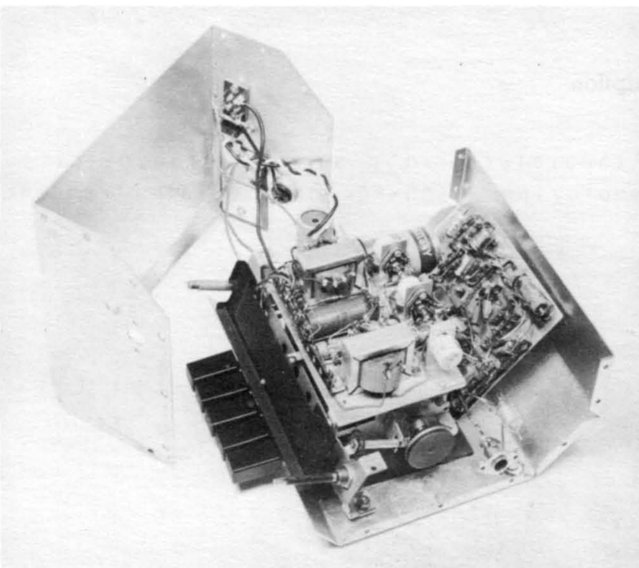


Fig. 2 - Receiver with cover removed.

*These data reflect the characteristics of the experimental transistors used; units of exceptionally high and exceptionally low performance were avoided. Long-term life performance of the transistors was not evaluated.

Circuit Description

A schematic diagram of the receiver is shown in Fig. 3. Transistors V1 through V6 are similar to the experimental units described in *LB-915, A P-N-P Triode Alloy Junction Transistor for Radio-Frequency Amplification*, and serve as r-f stage, mixer, oscillator, two i-f stages, and second detector. The audio complement, V7 through V9, are experimental p-n-p transistors electrically similar to those described in *LB-905, Power Junction Transistors by the Alloy Process*, but incorporate a mechanically and thermally improved mounting arrangement.

Tuner Transformers

The electrical elements of a push-button permeability tuner manufactured by Radio Condenser Corporation, Camden, N. J., were revised for operation in the transistor receiver. Three tightly-coupled transformers are employed in the antenna-to-rf input, rf-to-mixer interstage, and oscillator circuits. Maximum unloaded Q of these transformers is obtained by making the effective diameter of the windings as large as the tuner dimensions will permit. Tracking of the signal circuits is insured by maintaining equal effective diameters of the respective coils. Oscillator tracking is accomplished by approximating a variable-pitch oscillator-transformer winding, the desired variable pitch being obtained by changing the coil-winder gear ratio at intervals along the coil. The pertinent data for the antenna, interstage, and oscillator transformers are shown on the schematic.

The individual coil assemblies of the tuner are enclosed in shield cans. Magnetite sleeves of 3/8-inch inner diameter are provided for the antenna and interstage transformers. The powdered-iron tuning slugs are 1.2 inches long by 0.18 inch diameter; slug travel is about one inch.

Tuner Circuitry

A conventional automobile rod antenna is employed; this antenna feeds the tuned primary of the antenna transformer, T1, as shown on the schematic, Fig. 3. Since the antenna is effectively a voltage source in series with a capacitor, the power fed to T1 will increase with increasing Q and decreasing shunt capacitance. The shunt capacitance includes the capacitance of the shielded lead from the antenna, the stray capacitance of the wiring, and the reflected r-f stage input capacitance. The coil Q is limited by the winding dimensions to 50-70 across the band. The primary is designed to tune with a total of 75 μμf, allowing for the above shunt capacitance as well as the antenna capacitance and a trimmer capacitor, C1, of 3 to 30 μμf.

The choice of operating Q for the antenna and interstage transformers represents a compromise between the rejection of image and intermediate frequencies, which increases with operating Q, and insertion loss, which also depends on operating Q. Minimum insertion loss between the antenna and the r-f stage input obtains for the "matched" condition, i.e., when the turns ratio of T1 is adjusted so that the transformer tuned impedance, referred to the secondary winding, is equal to the input impedance of the r-f stage. For this condition, the operating Q is one-half the unloaded coil Q. For the compromise used, the operating Q is eight-tenths of the unloaded Q, with an associated increase in insertion loss of 2 db above that applying for the matched condition. This compromise is obtained by adjusting the turns ratio so that the secondary tuned-impedance is one-fourth the input impedance of V1. A typical value for the input impedance of V1 at midband is 75 ohms.

The r-f stage bias arrangement renders the stage relatively insensitive to changes in ambient temperature. The base is returned to a low resistance bias source of -1.5 volts at the junction of R2 and R3. Below a-g-c threshold, constant-emitter-current bias of 1.3 ma obtains, via the emitter resistor, R4, in conjunction with R5, R19 and R20. The emitter is returned to ground for r-f by C4. The bias conditions with respect to a-g-c action are discussed in connection with the second detector.

The r-f interstage transformer, T2, provides coupling from the collector of V1 to the base of the mixer, V2. In the mid-frequency range of the broadcast band the output impedance of the r-f transistor is 10,000 to 5,000 ohms, and the mixer input impedance is typically about 500 ohms. The operating Q of T2 is 15 to 20 and the transformer insertion loss is 3.7 db. This compromise between operating Q and insertion loss is obtained by adjusting the turns ratio of T2 so that the tuned primary impedance of T2 and the reflected input impedance of the mixer each equal the output impedance of V1. Tuning of the primary of T2 is provided by C6 and C7. The gain of the r-f stage is about 20 db at midband.

The collector circuit of V1 returns to the tap on the bleeder formed by R6 and R7. The bleeder tap is bypassed to ground by C5. This arrangement decouples the collector circuit from the common supply, and serves as a voltage divider to reduce the collector voltage of the r-f stage. An improved signal-to-noise ratio is obtained by operation at reduced collector voltage, as described in *LB-915, A P-N-P Triode Alloy Junction Transistor for Radio-Frequency Amplification.*

The tuned primary of the oscillator transformer, T3, in the collector circuit of the oscillator transistor, V3, affords an unloaded Q of from 40 to 60 over the oscillation frequency range. A relatively high tank capacitance is employed for stability. The number of secondary turns was determined experimentally for adequate mixer injection. The secondary applies feedback to the base of V3 and is returned to the -1.5 volt bias source at the junction of R2 and R3. The R10-C11 network in the emitter circuit of V3 introduces degeneration which reduces the net positive feedback in the oscillator circuit. The loading of the oscillator tuned circuit by the oscillator transistor input circuit is thus reduced so that oscillator tuning becomes relatively independent of the oscillator transistor input impedance. The reactance of the effective transistor base-to-emitter capacitance is in series with the relatively high reactance of C11. Thus, variation of the effective transistor base-to-emitter capacitance with frequency does not deteriorate oscillator tracking. (A decrease of the effective transistor base-to-emitter capacitance with increasing frequency arises from the presence of transistor base-lead resistance, which is in series with the emitter-junction capacitance. In the 1 to 2 Mc range, the base-lead resistance and the reactance of the emitter-junction capacitance are comparable in magnitude.) The resistor R10, in conjunction with the base bias, provides sufficient starting emitter current to initiate oscillation.

The secondary of the interstage transformer, T2, is coupled to the base of the mixer, V2. The emitter is returned to ground by R8, which provides bias stability in a manner analogous to that provided by the emitter-return resistors of the amplifier stages. Approximately 0.4 volt rms of oscillator injection is applied to the emitter through capacitor C8; the corresponding average emitter current is 0.4 ma. The optimum magnitude of oscillator injection depends in part on the magnitude of the emitter-return resistor; in this instance optimum injection is typically 0.35 volt rms. If injection decreases below this value, the conversion gain falls rapidly, while the conversion gain decreases relatively slowly with increasing injection. Somewhat greater than optimum injection insures interchangeability of mixer transistors and minimizes variation of conversion gain with small changes in oscillator injection.

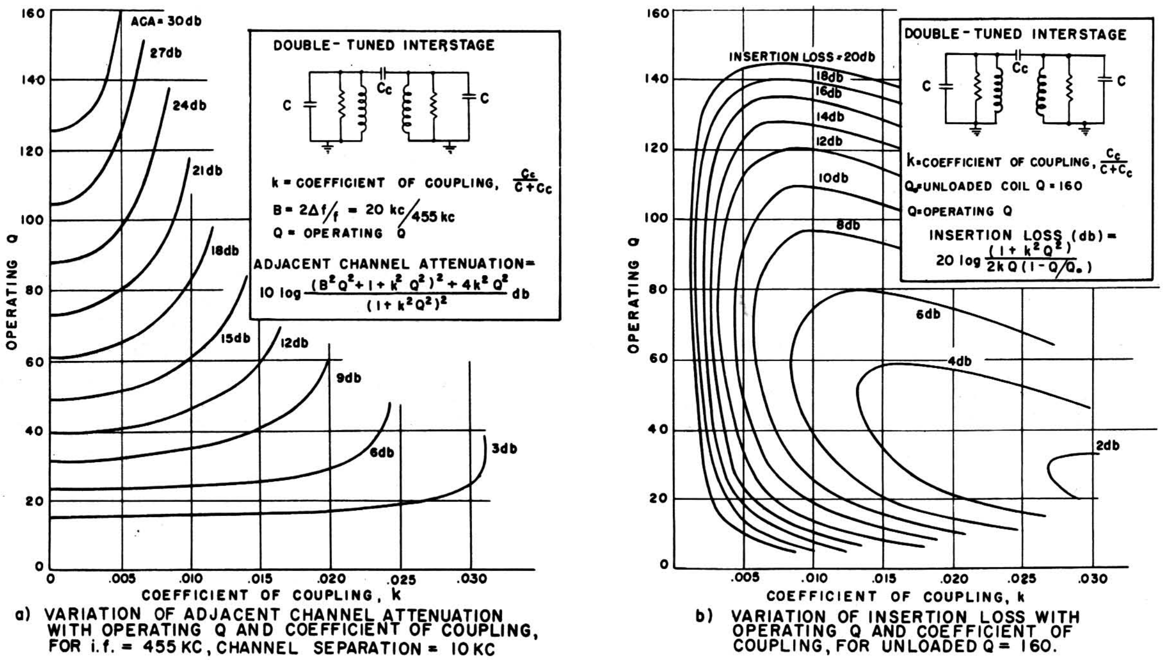
Since the coupling capacitor, C8, presents a low impedance at the intermediate frequency and the secondary of T3 has a low impedance at both signal and intermediate frequencies, R8 is effectively bypassed to ground for both the mixer input and output signals.

The mixer output impedance is typically 60,000 ohms and is relatively independent of signal frequency, while the input impedance is about 500 ohms at midband and decreases somewhat with increasing frequency. The mixer stage conversion gain is about 20 db at midband.

IF Amplifier

The i-f, detector, and a-g-c circuits are similar to those described in *LB-957, A Developmental Pocket-Size Broadcast Receiver Employing Transistors*. The three i-f interstage coupling networks consist of two capacitively-coupled double-tuned transformers, T4-T4 and T5-T5 and a single-tuned transformer T6. The mechanical arrangement of these transformers is described in *LB-931, Miniature IF Transformers*. The i-f amplifier contributes 37 db of adjacent channel attenuation (ACA); the selectivity provided by the various interstage networks is so apportioned as to minimize overall insertion loss for this ACA. The choice of turns ratio of a single-tuned transformer for minimum insertion loss at a prescribed ACA is described in *LB-957*.

For the double-tuned circuits, the ACA is determined by the coefficient of coupling, k, as well as the operating Q's, as shown in Fig. 4a. The insertion loss of the double-tuned circuits is determined by k, by the operating Q's, and by the unloaded coil Q, as shown in Fig. 4b. That optimum combination of k and operating Q which results in minimum insertion loss for a prescribed ACA may be determined from these curves. The variation of minimum



a) VARIATION OF ADJACENT CHANNEL ATTENUATION WITH OPERATING Q AND COEFFICIENT OF COUPLING, FOR i.f. = 455 KC, CHANNEL SEPARATION = 10 KC

b) VARIATION OF INSERTION LOSS WITH OPERATING Q AND COEFFICIENT OF COUPLING, FOR UNLOADED Q = 160.

Fig. 4 - Design curves for double-tuned i-f coupling transformers.

insertion loss vs ACA is shown in Fig. 5; a corresponding curve for a single-tuned circuit is plotted for comparison.

The ACA's provided by T4-T4, T5-T5, and T6 are 15.5 db, 15.5 db, and 6 db respectively, and the insertion losses are 5 db, 5 db, and 2.5 db respectively. The amplifier provides about 50 db gain from the base of the first i-f amplifier, V4, to the base of the second detector, V6.*

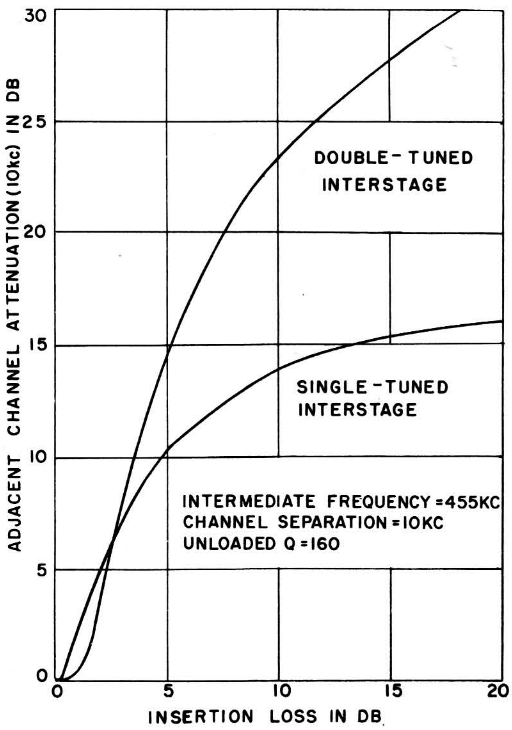


Fig. 5 - Minimum insertion loss vs ACA for double- and single-tuned circuits.

Biasing of the first i-f stage, to which a.g.c. is applied, is similar to that of the r-f stage. The second i-f stage is constant-emitter-current biased by means of the emitter resistor, R17. The i-f stage bases return to the tap on a separate bleeder formed by R12 and R13. The bleeder tap is bypassed to ground by C17. This bleeder circuit, in combination with the collector-circuit decoupling of the second i-f stage provided by R16-C23, serves to isolate the i-f amplifier from the r-f and mixer stages. Neutralization of the i-f stages is provided by C19 and C24.

*By way of example, i-f transistors having $r_{bb'}$ = 75 ohms, α_{cb} = 20, $C_{b'c}$ = 11 μμf, and $C_{b'e}$ = 0.001 μf ("alpha cutoff" of 6 Mc) would give typical performance.

Second Detector and AGC

The operation of the second detector is similar to that described in *LB-957, A Developmental Pocket-Size Broadcast Receiver Employing Transistors*. As shown in the schematic, a.g.c. is applied to the r-f and first i-f stages. Emitter current control is employed, and is obtained from the audio-agc detector, V6, via R20, R15, and R5. The bias arrangement of V1 and V4 holds the currents in R4 and R14 essentially constant; the controlled stages are constant-emitter-current biased for the zero-signal condition. An increase in the detector collector current produces a proportionate decrease in the emitter currents of the controlled stages. The magnitude of these currents at zero signal (about 1.5 ma per stage) is sufficient that the initial decrease in current results in little change in gain, introducing the desired delay in a-g-c action. The control current at the flat portion of the a-g-c characteristic is 3 ma; an additional 1 ma of detector current flows through the shunt resistor, R19. The total detector current of 4 ma is then sufficiently large to minimize the effects of elevated-temperature saturation current in the detector transistor. In the design of an a-g-c circuit of this type, adjustment of the zero-signal emitter currents of the controlled stages provides a convenient level control of the flat portion of the a-g-c characteristic.

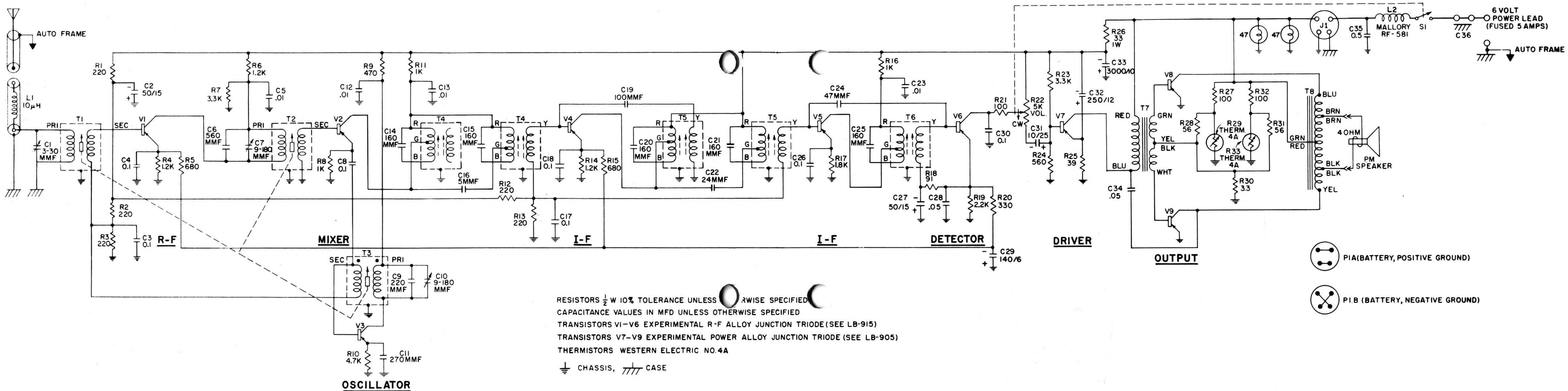
The R-C networks in the emitter circuits of the detector, r-f, and first i-f amplifiers provide filtering of the r-f and audio components of a-g-c current generated in the detector.

Detector linearity is improved by degeneration provided by R18, in the return path for the audio component of detector emitter current. Audio output is taken from the collector through the volume control, R22. Decoupling from the collector supply is provided by R21 and C30.

Audio Amplifier

The audio output of the detector is applied, via C31, to the base of the audio driver, V7. Since the dynamic output impedance of the detector transistor is high, the output impedance of the detector stage is essentially the resistance of the volume control irrespec-

An Experimental Automobile Receiver Employing Transistors

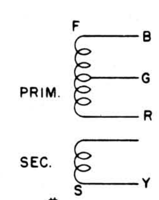


RESISTORS 1/2 W 10% TOLERANCE UNLESS OTHERWISE SPECIFIED
CAPACITANCE VALUES IN MFD UNLESS OTHERWISE SPECIFIED
TRANSISTORS V1–V6 EXPERIMENTAL R-F ALLOY JUNCTION TRIODE (SEE LB-915)
TRANSISTORS V7–V9 EXPERIMENTAL POWER ALLOY JUNCTION TRIODE (SEE LB-905)
THERMISTORS WESTERN ELECTRIC NO. 4A
CHASSIS, CASE

UNIT	PRIMARY		SECONDARY		FORM		GEAR SETTINGS		
	T	LITZ	T	SSE	I.D.	O.D.	FRONT	CAM	REAR
T1	322	4-43	5	#35	.187	.205	43-40	.125	28-100
T2	107	5-42	28	#35	.187	.205	50-40	.125	88-100
T3	102*	7-41	14	#35	.187	.250	43-40	.125	
	20								52-60
	25								46-60
	20								72-60
	20								65-60
	17								71-60

WOUND ON MEISSNER MACHINE - PRIMARY PROGRESSIVE UNIVERSAL ON BAKELITE FORM. SECONDARY DISTRIBUTED HAND WOUND OVER PRIMARY. *THE PRIMARY OF T3 IS A STEPPED-PITCH WINDING; THE STEPS ARE AS SHOWN

TUNER TRANSFORMERS

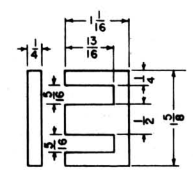


UNIT	SEC	PRIM		
		R	G	B
T4	4T	0T	55T	110T
T5	4T	0T	32T	110T
T6	14T	0T	40T	110T

#38 DSE ON 0.125 I.D. x 0.145 O.D. FIBRE FORM, RANDOM WOUND ON 0.230 WIDE x 0.125 D. ARBOR. SECONDARY WOUND UNDER PRIMARY - DISTRIBUTED OVER LENGTH OF COIL FINISHED COIL CUMAR DOPED.

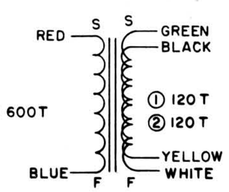
TRANSFORMER CORES AND MECHANICAL ASSEMBLY OF MINATURE I-F TRANSFORMERS DESCRIBED IN LB-931

I-F TRANSFORMERS



CORE: $\frac{1}{2}$" STACK .014 SILICON-IRON, LAP JOINT
WINDING LENGTH: $\frac{5}{8}$", .015 MANDREL

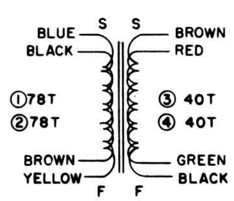
T7



WINDINGS ①, ② ARE BIFILAR:
120 TURNS #26HF, 15T/L, 8 LAYERS
.002 PAPER BETWEEN LAYERS
.004 PAPER BETWEEN ①, ② AND ③

WINDING ③:
600 TURNS #32 HF, 60T/L, 10 LAYERS
.001 PAPER BETWEEN LAYERS

T8



WINDINGS 1, 2 ARE BIFILAR:
78 TURNS #24 HF, 13T/L, 6 LAYERS
.002 PAPER BETWEEN LAYERS
.004 PAPER BETWEEN ①-② AND ③-④

WINDINGS 3, 4 ARE BIFILAR:
40 TURNS #22 HF, 10 T/L, 4 LAYERS
.002 PAPER BETWEEN LAYERS

AUDIO TRANSFORMERS

Fig. 3 – Schematic of transistor automobile receiver.

tive of volume control setting. The current available for driving the following stage is therefore proportional to the resistance intercepted between the slider and the common supply. Nearly all of this available output current flows into the base of the driver, since the driver presents an input impedance on the order of 25 ohms. A base-to-collector signal current gain of 40 to 50 is realized. For medium audio frequencies, the emitter of the driver is bypassed, by C32, to the common supply. The low-frequency response of the amplifier is down 3 db at that frequency at which the input impedance at the base of the driver (the product of the driver current gain and the reactance of C32) equals the parallel resistance of the volume control, R22, and the base-bias bleeder resistors, R23 and R24. The roll-off frequency, for operation into a 4-ohm dummy load, is approximately 120 cps. The response into the speaker load remains elevated to below 80 cps due to the mechanical low-frequency resonance of the loudspeaker.

Control of the driver stage operating current is provided by the base-bias bleeder in conjunction with the emitter resistor R25. The driver collector current is approximately 15 ma at moderate temperatures; it increases to 30 ma at 80 degrees C and drops to 10 ma at -40 degrees C.

The driver collector is transformer-coupled to the bases of the push-pull output stage by transformer T7. The interstage and output transformer data are given on the schematic.

In addition to the usual transformer design compromises, (efficiency, low-frequency response, etc.) the following requirements are met in the design of the interstage transformer:

1. The impedance reflected to the driver collector is low enough that driver overload does not occur before overload of the output stage.

2. The series resistance of the primary winding is low enough that the loss of driver supply voltage is tolerable.

3. The two halves of the secondary winding are sufficiently tightly coupled together to avoid transient voltages when current shifts from one output transistor to the other. This is accomplished by bifilar winding of the secondary.

A current gain of 5 is afforded by the interstage transformer from the collector of the driver to the bases of the output transistors. Non-linearity in the cross-over interval is minimized by providing the output transistors with a small initial threshold bias. Experiment indicates that there is an optimum value of threshold emitter current, about 20 ma, which results in minimum non-linearity. While this optimum value of current is essentially independent of temperature, the corresponding required base-to-emitter bias voltage varies with temperature at a rate of approximately -0.0025 volt per degree C. The resistor-thermistor networks, R27 through R33, provide a low-impedance bias source exhibiting the proper variation of voltage with temperature. The transistor-thermistor mounting arrangement, shown in Fig. 6, provides for close thermal contact between the transistor case and mounting bracket, and thermistor, mounting bracket, and chassis.

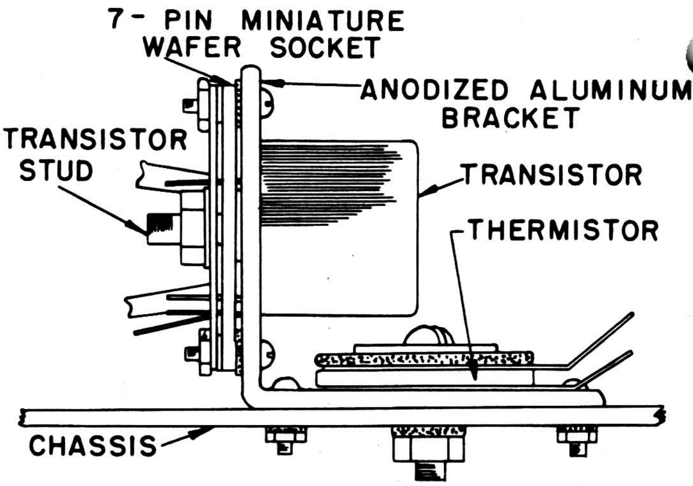


Fig. 6 - Power-transistor and thermistor mounting arrangement.

The output stage affords large-signal current gain of 20 to 30, and transconductance in excess of 1 mho.* With a 6.6-volt collector supply-voltage, a collector voltage swing of 6 volts peak is obtainable. The corresponding peak collector current, for 2.0 watts output (10 per cent distortion), is 0.67 ampere. The

*These characteristics were obtained with experimental transistors having substantially better high-current performance than that shown in *LB-905*.

4-ohm speaker voice-coil impedance is transformed to the required 9-ohm collector load by transformer T8. The use here of a bifilar wound autotransformer provides close coupling and high transformer efficiency. The power gain of the output stage is on the order of 24 db.

High-frequency roll-off of the audio amplifier is controlled by capacitor C34, which introduces inverse feedback in the output stage at frequencies above 2 kc.

The audio input circuits are decoupled from the power supply by R26 and C33.

Power and Interference Considerations

Conventional means are utilized to eliminate high-frequency interference on the power lead. A spark plate, C36, and an r-f choke and capacitor L2, C35, remove the VHF and MF components respectively. The receiver chassis is insulated from the receiver case. Power connection for operation in automobiles with positive or negative polarity battery ground is obtained by employing plug P1A or P1B, respectively.

Rejection of high-frequency impulse type of ignition interference appearing on the antenna is accomplished by the choke, L1, in series with the antenna lead, which, together with the shunt capacitance across the antenna primary, forms a low-pass filter.

An additional potential source of interference arises from the periodic current drawn by the automobile ignition system, which may produce a low-frequency voltage fluctuation on the power lead. The presence of more than a few millivolts of this type cf interference appearing between bases and emitters of the gain-controlled stages would produce objectionable modulation of the received signal. This is avoided by returning the bias bleeders, as well as the collectors of V1 through V6, to the decoupled voltage available at the junction of R26 and C33, and by the additional isolation of the base returns provided by C2.

Performance

Various receiver performance characteristics are shown in Fig. 7.

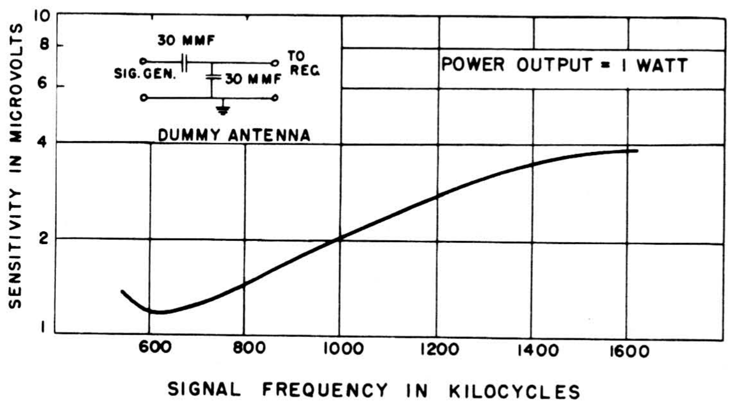
The receiver sensitivity as a function of signal frequency appears in Fig. 7a. The increase in sensitivity at the lower extreme of the broadcast band is due largely to the increased gain of the r-f and mixer stages over that obtaining at the upper extreme of the band. The shape of this curve is also influenced by the tracking error of the oscillator transformer which is plus-or-minus 4 kc. This tracking error is due in part to the step approximation to a variable-pitch winding employed in this transformer.

The i-f and overall selectivity at 1000 kc, and the i-f and image rejection as a function of receiver tuning are shown in Figs. 7b and 7c respectively. These curves reflect the design compromises in the antenna, interstage, and i-f coupling networks. The overall adjacent channel attenuation at 1000 kc is 41 db. Since the overall selectivity is determined almost entirely by the i-f coupling networks, the ACA is essentially independent of the signal frequency. The image rejection at 1600 kc is 42 db and increases to about 60 db at the low end of the band. The i-f rejection provided at 540 kc is 40 db and increases rapidly with increasing signal frequency.

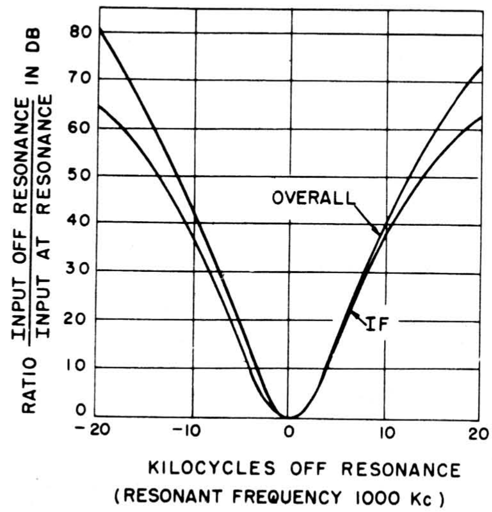
The a-g-c and noise characteristics are shown in Fig. 7d. The a-g-c figure of merit is 63 db. The signal input required for a 20-db signal-to-noise ratio is 12 μvolts; the corresponding equivalent-noise-sideband input is 0.4 μvolt.

The distortion vs power-output characteristic is displayed in Fig. 7e. Total harmonic distortion reaches 10 per cent, mostly third harmonic, at approximately 2 watts output. At this level, distortion arises almost entirely from the curvature of the current-gain characteristics of the output transistors. Since the power output is not limited by the dissipation capabilities of the transistors, operation in automobiles with 12-volt electrical systems should permit a substantial increase of power output, utilizing the same transistors at the same peak currents.

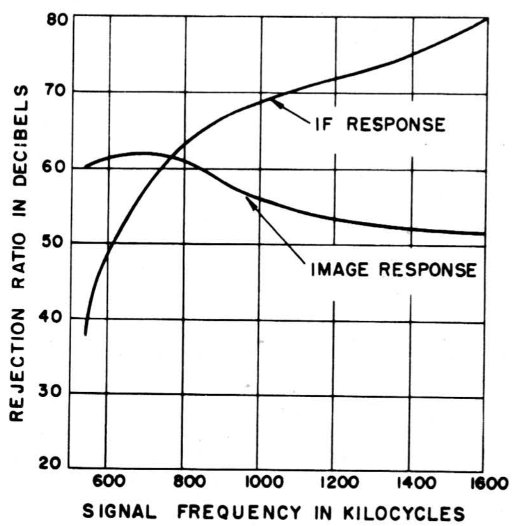
The distortion vs per cent modulation characteristics for 1.0 watt and 100 milliwatts output are shown in Fig. 7f. The distortion is substantially constant up to 50 per cent modulation and increases above this level due to an increasing detector contribution.



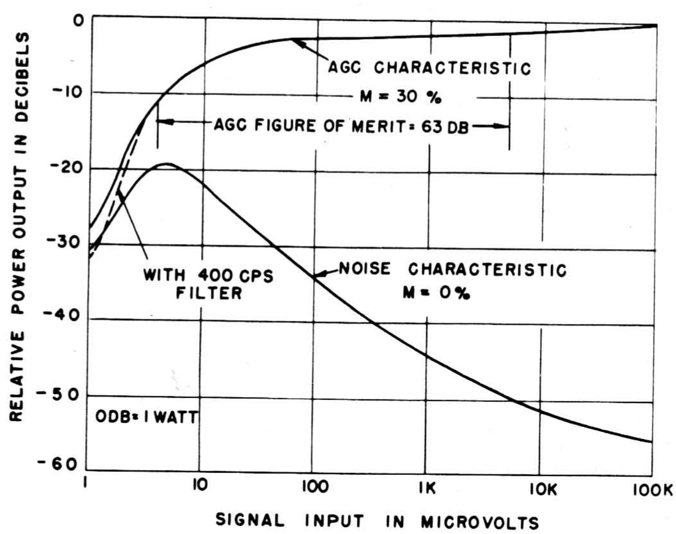
a) SENSITIVITY VS SIGNAL FREQUENCY



b) IF AND OVERALL SELECTIVITY

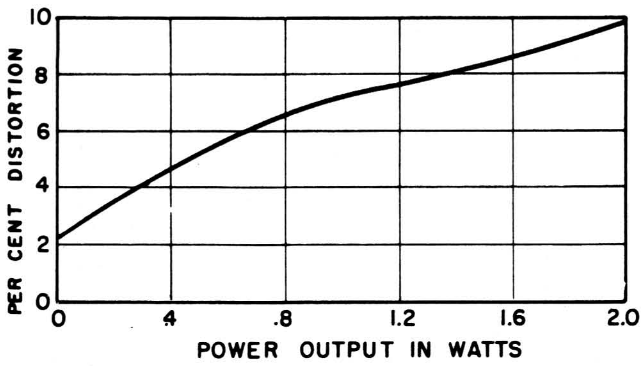


c) IF AND IMAGE REJECTION CHARACTERISTICS.

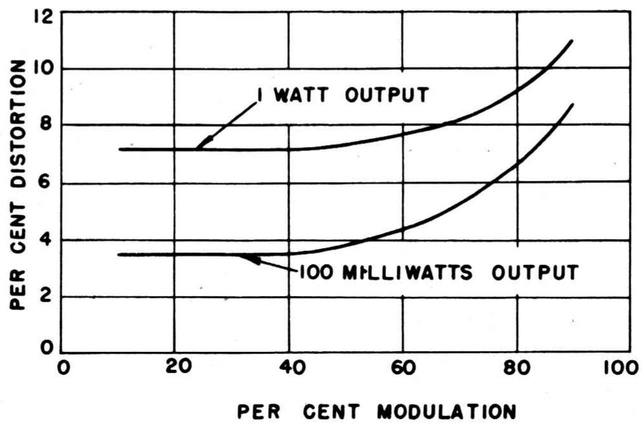


d) AGC AND NOISE CHARACTERISTICS

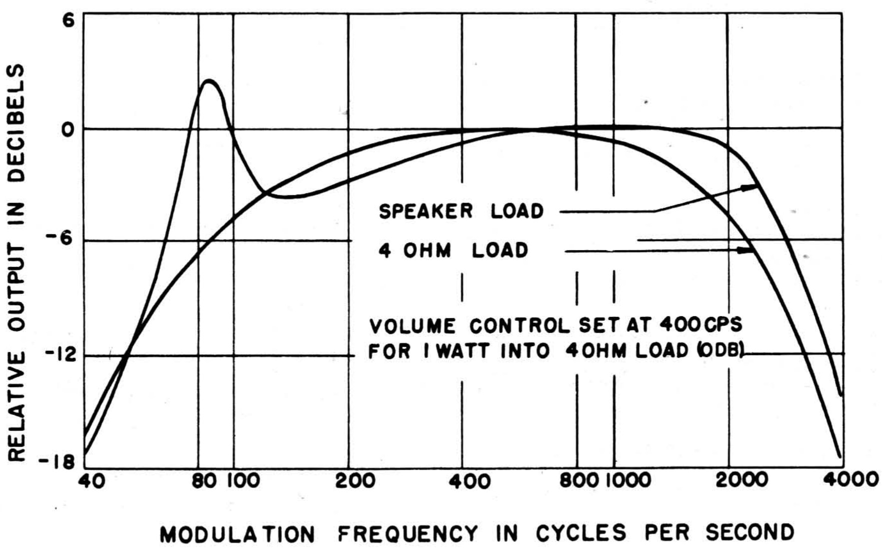
Fig. 7 – Receiver performance characteristics.



e) DISTORTION VS POWER OUTPUT



f) DISTORTION VS PER CENT MODULATION



g) ELECTRIC FIDELITY

Fig. 7 – Receiver performance characteristics.

The electric fidelity into the speaker load and into a 4-ohm dummy load is displayed in Fig. 7g. The mechanical resonance of the 6x9-inch speaker and the series inductance of the voice coil account for the elevated response into the speaker-load in the vicinity of 85 and 2000 cps respectively. Volume-control compensation has not been employed; the electric fidelity is essentially independent of volume-control setting. The receiver acoustical performance is influenced by the effective baffle which is employed, and thus is different for various makes and models of automobiles.

Curves illustrating the performance of the receiver as a function of ambient temperature are shown in Fig. 8.

Fig. 8a shows the receiver sensitivity and the signal required for 20-db signal-to-noise ratio as a function of ambient temperature. The sensitivity is 2.0 µvolts at 20 degrees C, and is below 10 µvolts over the range from -40 degrees C to 80 degrees C. Loss of sensitivity at high temperatures arises from the reduced Q of the core materials in the tuned circuits, from detuning effects, and from a decrease in gain of the high-frequency transistors of approximately 1 db per stage. The receiver noise performance is substantially unaffected over this range of temperature.

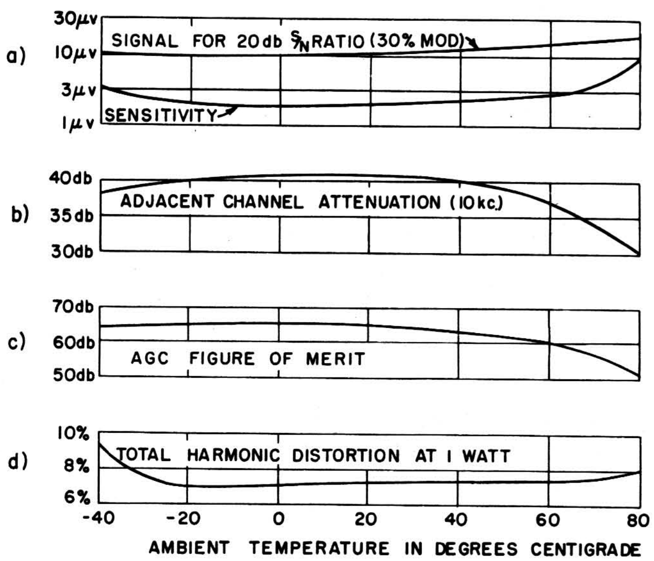


Fig. 8 - Receiver performance as a function of ambient temperature.

Fig. 8b shows the receiver ACA as a function of ambient temperature. Differences in tuned-frequency shift of the individual i-f interstages tend to reduce selectivity at the extreme temperatures. This tendency is offset by the increase in Q's at reduced temperatures, and aggravated by the reduction in Q's at elevated temperatures, resulting in a reduction in ACA from approximately 40 db, at low and moderate temperatures, to 31 db at 80 degrees C.

The variation of the a-g-c figure of merit with ambient temperature is shown in Fig. 8c. The figure of merit is approximately 65 db at low and moderate temperatures, and drops to 51 db at 80 degrees C. The shape of the a-g-c characteristic at any temperature in this range is essentially the same as that shown in Fig. 7d. The reduction in a-g-c figure of merit at elevated temperatures arises from the corresponding reduction in receiver sensitivity.

The total harmonic distortion at 1.0 watt output, shown in Fig. 8d, and the maximum power output (10 per cent distortion), are relatively unaffected over the temperature range from -40 degrees C to 80 degrees C. The increase in distortion at extreme temperatures arises for the most part from the imperfect temperature compensation provided by the output-stage biasing network. Additional distortion at -40 degrees C originates in the detector-agc circuitry, and is due mainly to the increased impedance of the sintered tantalum electrolytic capacitor, C29, in the a-g-c filter network. It is the use of this type of capacitor here, notable for good low-temperature performance, that permits operation to -40 degrees C. Substitution of a conventional aluminum-foil type capacitor restricts the useful lower-temperature limit to approximately -10 degrees C.

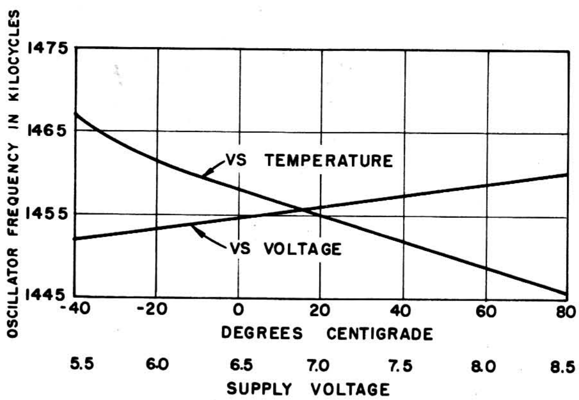


Fig. 9 - Oscillator frequency stability.

The change in oscillation frequency with change in supply voltage, shown in Fig. 9, is about 0.2 per cent per volt. The frequency variation of the oscillator with respect to ambient temperature is also shown in Fig. 9. The difference in oscillator-frequency shift and in the frequency shift of the signal tuned-circuits is in the same direction and of approximately the same magnitude as the shift in the i-f tuned-circuits. The net tracking error as a function of ambient temperature is therefore smaller than either the oscillator- or intermediate-frequency shifts.

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